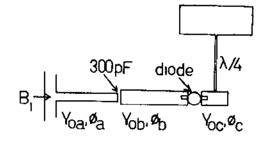

Fig. 3. Top view of one of the stripline stubs.

TABLE I
STUB DESIGN PARAMETERS FOR REALIZABLE PHASE SHIFTER IN FIG. 3

	$\theta = 90^\circ$		$\theta = 75^\circ$	
	ϕ	Y_0, mhos	ϕ	Y_0, mhos
A	76.24°	0.0105	102.7°	0.00986
B	30°	0.020	30°	0.020
C	20°	0.020	20°	0.020

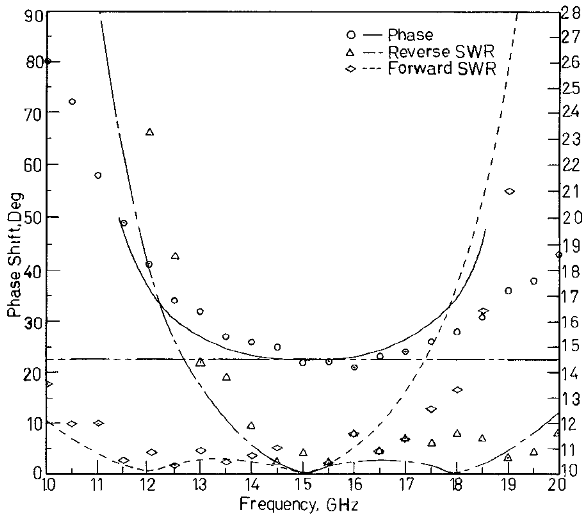

Fig. 4. The theoretical and experimental phase shift and SWR when the susceptance spacing is 90°.

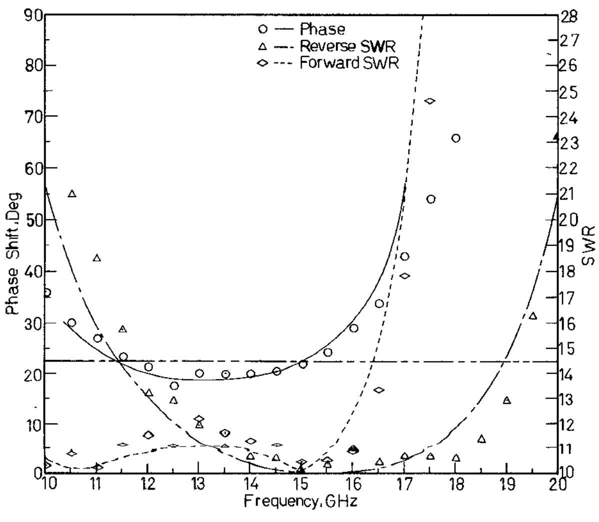

Fig. 5. The theoretical and experimental phase shift and SWR when the susceptance spacing is 75°.

drops from 43 percent, when $\theta = 90^\circ$, to 12 percent, when $\theta = 75^\circ$. Clearly, the 90° susceptance spacing phase shifter has superior bandwidth properties.

A test was made with a practical circuit to see if the bandwidth decreases as the susceptance spacing is reduced from 90°. Two 22.5° phase shifters were designed, analyzed, and built in stripline: one with $\theta = 90^\circ$ and the second with $\theta = 75^\circ$.

The first step in the design was to measure the impedance parameters of two shunt mounted pin diodes in the two bias states. To get the B_i required by (8) and (9), each diode was mounted in a stub circuit as shown in Fig. 3. Calculations for several stub configurations were made, each giving the required B_i, but one was chosen that had reasonably short line lengths and realizable characteristic admittances. Table I shows the design parameters used for the stubs, while Figs. 4 and 5 show the experimental and theoretical SWR and phase shift for the two phase shifters. Here the bandwidth decreased from 19.3 percent for $\theta = 90^\circ$ to 5.75 percent for $\theta = 75^\circ$. For both this circuit and the one using lumped reactances the bandwidth was reduced by a factor of approximately 3.5.

CONCLUSIONS

The commonly used spacing of $\theta = 90^\circ$ between shunt susceptances appears to offer the widest bandwidth available for a loaded-line phase shifter. Shortening θ to, say, 75°, implies that one of the required $|B_i|$ are larger than the $|B_i|$ when $\theta = 90^\circ$. Any bandwidth advantage gained through the shorter line length θ is overcome by the larger required shunt susceptance.

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Circularly Polarized Electric Field in Rectangular Waveguide

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***Abstract*—The electric field in a waveguide partially filled with a low-loss slab of dielectric in the *H* plane presents a circularly polarized component at the air–dielectric interface over a limited frequency range. This effect could be used to improve the performance of nonreciprocal devices utilizing the gyroelectric effect in magnetized semiconductors.**

I. INTRODUCTION

Nonreciprocal wave propagation has been observed in rectangular waveguides loaded with semiconductor slabs at room temperature subjected to a dc-biasing magnetic field. Both E- and H-plane structures were tested, using, respectively, N-type InSb [1] and N-type silicon [2]. Devices such as circulators, isolators, and nonreciprocal

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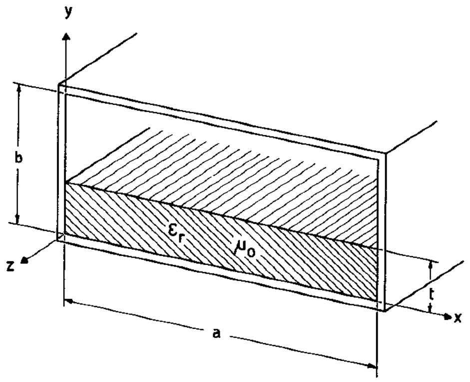

Fig. 1. Rectangular waveguide loaded with an H-plane slab next to the guide wall.

phase shifters could in principle be realized using this effect at millimeter-wave frequencies. However, a rather important drawback is the large attenuation reported so far for room-temperature operation. In gyroelectric semiconductors, nonreciprocity is due to the anisotropic conductivity tensor, which is inherently lossy. On the other hand, the nonreciprocal effects in gyromagnetic ferrites are due to an anisotropic permeability tensor, which is not directly connected with lossy behavior, except in the vicinity of the resonant frequency (very low losses can be achieved in nonreciprocal ferrite devices operating far from the resonance).

Lower losses in semiconductor devices could be obtained by decreasing the conductivity of the material used. This would, however, not be a practical solution, as it would at the same time significantly reduce the desired nonreciprocity. An improvement could be obtained by reducing the amount of semiconducting material (thin film) and placing it in a position where its nonreciprocal effect would be maximum, i.e., in a plane of circularly polarized *electric* field.

It is well known that the dominant TE_{10} mode in an empty rectangular waveguide possesses two planes of circularly polarized magnetic field [3]. Locating magnetized ferrite slabs at or close to these planes allows one to optimize the performance of nonreciprocal devices [4]. Loading with dielectric slabs also plays an important role. It is suggested here that a similar optimization is possible for semiconductor devices by placing the gyroelectric material at a plane of circularly polarized electric field. Such a plane does not exist in empty guide; however, once the guide is loaded with an H-plane dielectric slab next to the broad wall (Fig. 1), the electric field at the interface presents a circularly polarized component perpendicular to the $\mathbf{a}_x$ axis over a certain frequency range. A thin layer of low-loss semiconductor material located on the interface will then provide an optimal reciprocity to loss ratio when subjected to a biasing magnetic field in the $\mathbf{a}_x$ direction.

II. THEORY

The study of electromagnetic-field propagation in a rectangular waveguide containing H-plane dielectric slabs is already available in the literature: the propagation constant for the dominant LSM_{11} (quasi-TE_{10}) mode was determined in the case of lossless dielectrics in [5], which also considered the frequency limitations due to higher order modes. More recently, the study was extended to lossy dielectrics and the field distributions were also determined [6]. For lossless or low-loss dielectric loading, the E_z and E_y components of the electric field are in quadrature. Circular polarization then occurs when the amplitudes are equal. The amplitude ratio of these two components at the interface was determined by means of the computer program developed in [6]. The ratio within a thin slab of semiconductor can then be determined from consideration of the boundary conditions, assuming that this slab does not perturb excessively the electromagnetic-field distribution.

The tangential electric field (E_z component) is continuous:

$$E_{zd} = E_{zs} \tag{1}$$

where the index d indicates the field in the dielectric and s that in the semiconductor.

The normal component E_y of the field satisfies the following relation, in which ρ_s is the surface charge on the interface:

$$\epsilon_d E_{yd} - \epsilon_s E_{ys} = \rho_s \tag{2}$$

where ϵ_d and ϵ_s are, respectively, the relative permittivity of the dielectric and of the semiconductor.

The continuity relation for the current density (which vanishes in the lossless dielectric but is nonzero in the semiconductor) then allows one to determine the surface charge

$$-\sigma_s E_{ys} = -j\omega\rho_s. \tag{3}$$

Solving for the ratio E_y/E_z we then obtain

$$\left.\frac{E_y}{E_z}\right|_{\text{semiconductor}} = \frac{\epsilon_s + \sigma_s/j\omega}{\epsilon_d} \left.\frac{E_y}{E_z}\right|_{\text{dielectric}} \tag{4}$$

The electric-field components E_y and E_z will remain in phase quadrature only when the ratio $\sigma_s/\omega\epsilon_s$ is small, i.e., for small conductivity and high frequency.

III. CALCULATED RESULTS

A rectangular waveguide having an aspect ratio 2:1 was considered in the calculations. It contains a semiconductor of relative permittivity $\epsilon_s = 16$, placed on top of a dielectric slab of same permittivity (Fig. 2), and on a dielectric slab having a larger permittivity $\epsilon_d = 50$ (Fig. 3). Circular polarization occurs when $E_y/E_z = 1$ (solid line on drawings), which happens only at one frequency for a given dielectric thickness. The area over which the opposite circularly polarized component is 20 dB or more below the main one is delimited by the two dotted lines. It is thus possible to define a frequency bandwidth for operation, which varies from about 5 percent for the thicker slabs to about 20 percent for the thinner ones. Loading with

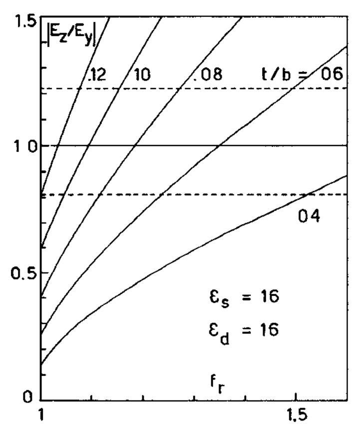

Fig. 2. Frequency dependence of the ratio $|E_z/E_y|$ in a thin semiconductor slab placed on top of a lossless dielectric having the same permittivity (f_r is the frequency normalized with respect to the cutoff of the dominant mode in the empty guide).

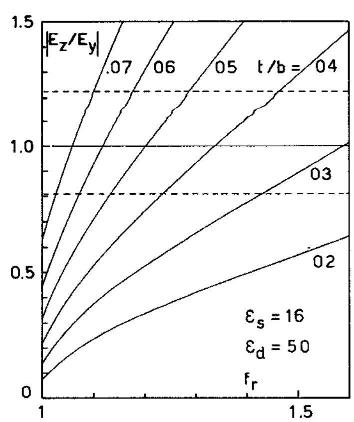

Fig. 3. Frequency dependence of the ratio $|E_z/E_y|$ in a thin semiconductor slab placed on top of a lossless dielectric having a larger permittivity.

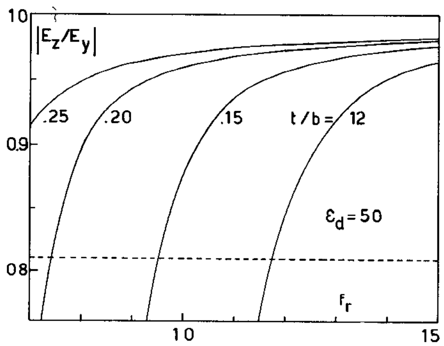

Fig. 4. Frequency dependence of the ratio $|E_z/E_y|$ in air over a high-permittivity lossless dielectric.

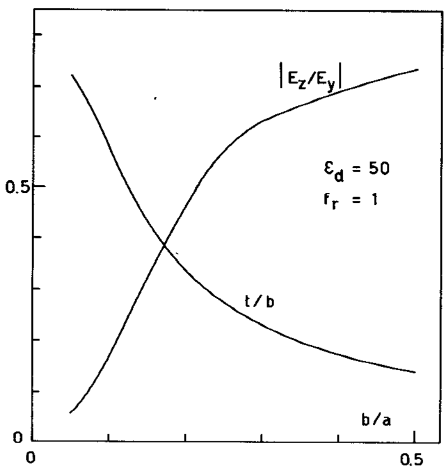

Fig. 5. Effect of the waveguide aspect ratio b/a on the maximum allowable filling factor t/b at the LSM_{21} cutoff and corresponding (maximum) ratio $|E_z/E_y|$.

high-permittivity dielectric does not increase the bandwidth. The waveguide loading needed to provide circular polarization is small enough to ensure single-mode operation [5].

When thick high-permittivity slabs are utilized, the electric field in the air next to the dielectric is almost circularly polarized over a much larger frequency range (Fig. 4). This effect could be quite useful for experiments involving thin gaseous plasmas. Unfortunately, it occurs at frequencies larger than the cutoff of higher order modes, hence particular precautions would be needed to avoid their excitation. The possibility of varying the waveguide aspect ratio b/a in order to raise the cutoff frequency of the first higher order mode was considered. For the particular range of interest, this mode is the LSM_{21} mode (quasi-TE_{20}), the cutoff frequency of which occurs when [5]

$$\beta_{11} = \sqrt{3}\pi/a \simeq 5.44/a \tag{5}$$

where β_{11} is the phase coefficient of the propagating dominant LSM_{11} mode. Computations were carried out to determine the effect of the aspect ratio b/a. The reduction of height, keeping the same filling factor t/b, does actually raise the LSM_{21} cutoff frequency. Alternatively, at one given frequency, operation with a larger filling factor would be possible. Unfortunately, it also decreases the ratio E_z/E_y at the interface. This effect is illustrated in Fig. 5, which presents at one fixed frequency ($f_r = 1$) the filling factor t/b and the corresponding ratio E_z/E_y at the LSM_{21} cutoff as a function of aspect ratio b/a. It is quite clear that height reduction is not suited to avoiding higher order modes, as it affects quite adversely the occurence of circular polarization.

IV. CONCLUSION

Loading a rectangular waveguide with a thin slab of low-loss dielectric provides a simple means for obtaining circularly polarized electric fields. This effect could be used to reduce the losses of nonreciprocal semiconductor devices and to study gyroelectric properties in semiconductors and gaseous plasmas.

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A 40–120-GHz Michelson Interferometer-Type Band-Splitting Filter

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Abstract—**The experiments of a Michelson interferometer-type filter with a superbroad frequency range 40–120 GHz are described. The branching loss is 1.2–2.2 dB within this frequency range. The filter can be used for a 40–120-GHz millimeter-wave waveguide transmission system.**

A millimeter-wave waveguide transmission system uses a frequency range from 40 to 120 GHz and requires three kinds of band-splitting filters with different bandwidths, 20, 40, and 80 GHz, to split the band into eight sub-bands [1]. Several band-splitting filters have been proposed [2]–[5]. These filters have been verified for the bandwidth 20 or 40 GHz. But, in order to increase the bandwidth more than 40 GHz (80 GHz, say), a Michelson interferometer-type filter may only be the device that meets these requirements.

This filter was first reported by Marcatili and Bisbee and experimentally tested in frequency range from 40 to 80 GHz [5]. Iiguchi developed the design theory of the filter [6].

This short paper describes the more detailed experiments of the filter whose bandwidth is 80 GHz, and gives data useful enough to design for the practical use.

This filter consists of Michelson interferometer-type hybrid circuits which are based on the concept of quasi-optical technique, two cutoff filters with high-pass responses, and other components as shown in Fig. 1. The frequency band from f_L to f_H (f_L,f_H: the lowest and highest frequencies of the band, respectively) coming in at port W is split into two sub-bands, the lower band from f_L to f_C (f_C: cutoff frequency of the cutoff filter) which emerges from port L and the higher band from f_C to f_H which emerges from port H, respectively.

First, the wave in the lower band is not related to the hybrid H_B because it only round trips in the hybrid H_A. Therefore, the hybrid H_A can be designed such that the coupling deviation in the band

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